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Some considerations on the use of pre-amplifiers at low-power u.h.f. relay stations

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SOME CONSIDERATIONS ON THE USE OF PRE-AMPLIFIERS AT LOW-POWER UHF RELAY STATIONS

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Summary

A number of inexpensive, commercially available pre-amplifiers are examined and their performances in respect of gain, noise factor, linearity and impedance match are compared. Some theoretical planning curves are presented for considering low-field conditions and formulae are derived for dealing with high-signal conditions.

A re-broadcast link front-end, made up from the most suitable of the amplifiers examined, is assessed in respect of its overall performance. An indication is given of the amount of isolation which is required between transmitting and receiving aerials and the desirable response of the mast-head filter.

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(RA-153)

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List of symbols

f	frequency MHz
s	signal power level subscripts: i = input, o = output 1,2,3 identification
s'	relative power level, subscripts as for s
v	general voltage level subscripts: i = input, o = output
x	slope of filter band-edge, dB/MHz
B	bandwidth
C	output intercept point, dB(mW)
D	feeder loss
E	field strength, dB(μV/m)
F	noise factor, subscripts for specific values
G	power gain, subscripts for specific values
G_A	aerial gain (relative to a half-wave dipole)
G_α	amplifier power gain
I_A	aerial isolation, dB
i_E	emitter current
M	noise measure
N	noise power level
P,Q	standards of television transmission
R	resistance, subscripts for specific values
T	temperature, Kelvin
V	peak voltage level, subscripts for specific values
Y_s	source admittance
Z	constant
α,β	constants
λ	wavelength, m
ω	angular frequency

(RA-153)

1. Introduction

As the total number of u.h.f. television relay stations in the UK is likely to be several hundred, ways of reducing costs are continually being considered. A preliminary theoretical survey has indicated that a reasonable number of stations will require only one, or maybe two, log-periodic aerials for the receiving array. This is possible when co-channel interference conditions allow a broader horizontal radiation pattern to be used, e.g. when the terrain provides some protection. By using pre-amplifiers to compensate for reduced aerial gain, the number of sites at which the simpler receiving aerial is suitable can be increased.

A second benefit of using pre-amplifiers is the improvement in the signal-to-noise performance in low field conditions. The benefit is greatest when the amplifier is used at the top of the mast immediately after the aerial. The attenuation of the feeder then contributes little to the overall noise performance and a cheaper, higher-loss cable can be used.

As an alternative, an amplifier at the foot of the mast, immediately in front of the channel-splitting equipment and transposers, would also be beneficial provided its noise factor is better than that of the transposer. In this case a low-loss feeder is desirable.

The report first considers some of the theoretical aspects of using pre-amplifiers at u.h.f. relay stations. An assessment is then given of a number of commercially available inexpensive amplifiers. In addition a series of measurements has been made to establish the aerial isolation templet which is appropriate to low-power relay stations.

2. Theoretical considerations

2.1. Low signals

The sensitivity of a receiver, which determines the lowest acceptable signal levels, is related to its noise factor, F_s (or to its noise temperature).[1]

The video signal-to-noise ratio at the output of a system is numerically given by (see Appendices 1 and 2)

$$\frac{S}{N} = E + 20\log\lambda + G_A - F_s - 25{\cdot}7\ \text{dB} \qquad \text{(i)}$$

where E = received field strength, dB(μV/m)

λ = signal wavelength, m

G_A = net aerial gain, dB

F_s = receiving system noise factor, dB

If the required signal-noise ratio is specified at the output of the receiver, the lowest field strength required can be found from this expression in terms of the various parameters.

The transmission standards for u.h.f. low-power relay stations have already been established and depend on the type of station. Details of the two standards, which are referred to as P and Q, are given in Appendix 3. The minimal acceptable received field strength at a frequency of 650 MHz in relation to the system noise factor, is shown in Fig. 1. Two types of receiving aerial are assumed: one comprising a single log-periodic aerial and the other an array of four log-periodic aerials.

The system noise factor depends on the arrangement adopted, i.e. whether the pre-amplifier is placed at the mast-head immediately after the aerial, or at the foot of the mast following the main feeder.

Referring to Figs. 2(a) and (b) and assuming a matched system, the respective noise factors are given by:

$$F_s(a) = F_1 + \frac{1}{G_1}\left(\frac{F_2}{D} - 1\right) \qquad \text{(ii)}$$

$$F_s(b) = \frac{1}{D}\left(F_1 + \frac{1}{G_1}\left[F_2 - 1\right]\right) \qquad \text{(iii)}$$

where the parameters as listed below are expressed in power ratios

F_1, G_1 are the pre-amplifier noise factor and gain respectively
F_2 is the receiver noise factor, and
D is the feeder loss

The performance of these arrangements is compared with a conventional arrangement in Table 1. A receiver noise factor of 10 dB, a feeder loss of 4 dB, a pre-amp noise factor of 4 dB and a pre-amp gain of 15 dB, are assumed:

TABLE 1

Minimal received field strength in dB(μV/m) required for some r.b.l. systems using 4 log-periodic aerials at 650 MHz

	$\frac{S}{N}$=45 dB	$\frac{S}{N}$=39 dB(P)	$\frac{S}{N}$=33 dB(Q)
No pre-amp F_s = 14 dB	78·9	72·9	66·9
Mast-base amp F_s = 8 dB	72·9	66·9	60·9
Mast-head amp F_s = 5 dB	69·9	63·9	57·9

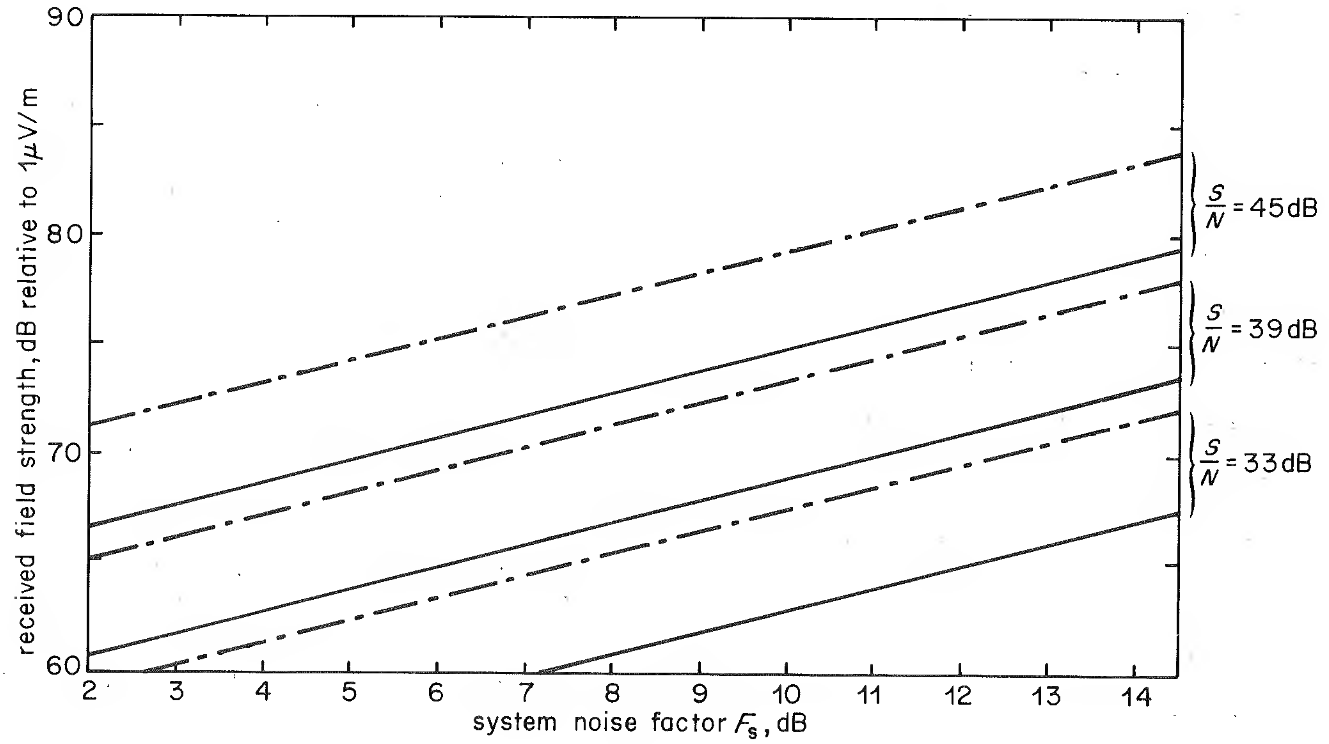

Fig. 1 - The dependence of the received field strength on system noise factor at 650 MHz

——— 4 log-periodic aerials   — · — Single log-periodic aerial

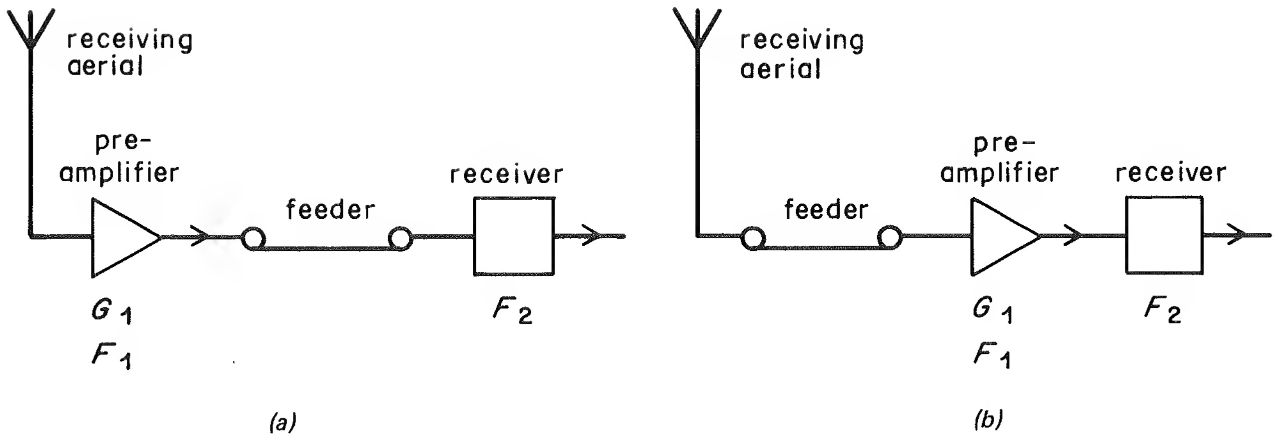

Fig. 2 - Receiving systems

(a) Mast-head pre-amplifier (b) Mast-base amplifier

The advantage of a pre-amplifier in low-signal conditions is readily seen; moreover, the advantage is greatest when a mast-head pre-amplifier is used. For this case, Fig. 3 shows how the system noise factor depends on the feeder loss and pre-amplifier gain. A maximum gain of 20 dB is seen to be adequate for this case. Only if the feeder loss were high or the receiver more noisy, would a greater gain be required.

2.2. High signals

At high signal levels an amplifier's non-linearity causes it to generate intermodulation products (i.p.s.) and eventually to overload. Let the amplifier characteristic be represented by a third order polynomial such that the output response, v_o, to an input signal, v_i, is

$$v_o = \alpha\, v_i + \beta\, v_i^3$$

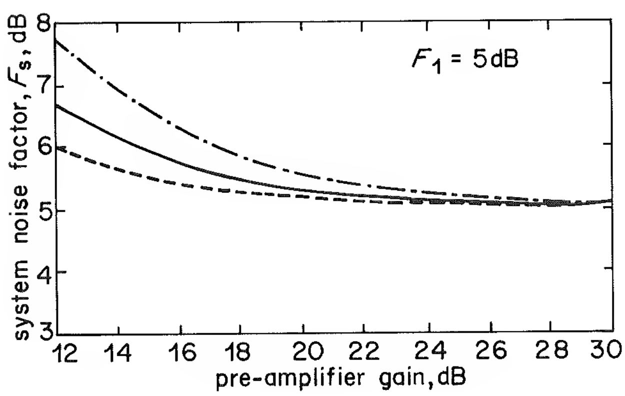

Fig. 3 - The noise factor for system in Fig. 2(a)

— —— — Feeder loss = 6 dB
———— Feeder loss = 4 dB
- - - - - - Feeder loss = 2 dB

where α and β are constants. For the signal excursions being considered, this representation gives a good approximation for calculations involving solid-state amplifiers.

Consider the effect of applying two spectral components to the amplifier:

$$v_i = V_r \cos \omega_r t + V_s \cos \omega_s t$$

The output spectrum in the neighbourhood of the input frequencies is of the form shown in Fig. 4, where the amplitude of the i.p.s. are ¾ $\beta . V_r^2 V_s$ and ¾ $\beta . V_r V_s^2$. Third order i.p.s. of this type have the largest amplitude when several television carriers, in a standard channel grouping, are applied to an amplifier.

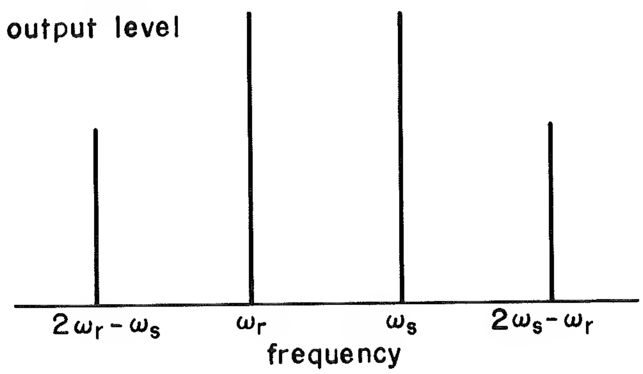

Fig. 4 - The output spectrum of a non-linear amplifier with two input tones

The principal third-order i.p.s. mentioned above can be used to define the linearity and hence the dynamic range of an amplifier. A "two-tone" test is applied in which two equal sinusoidal signals of different frequencies at the input are progressively increased in amplitude. The input/output characteristics of the power levels associated with the fundamental and i.p.s. are plotted on a graph in dB(mW) in Fig. 5.

The curves are linear for low input levels and their

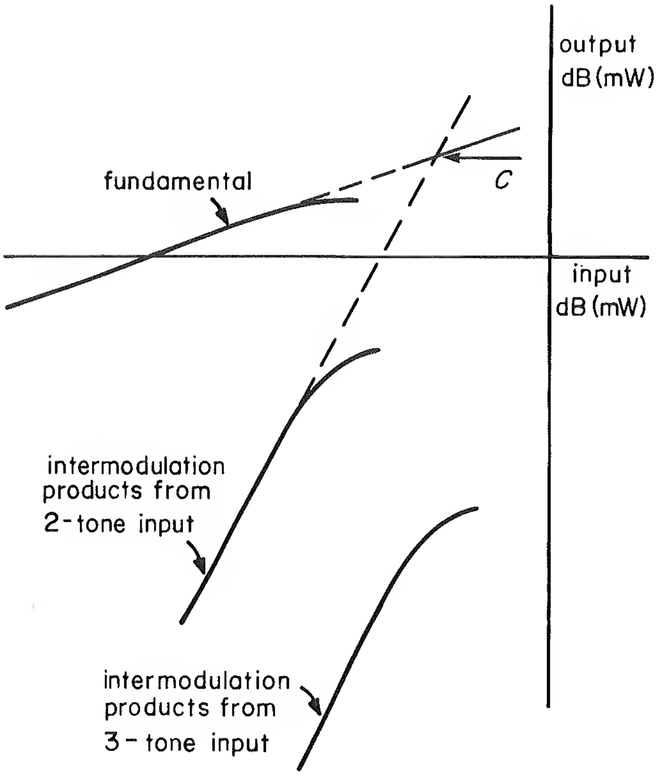

Fig. 5 - The output response of a non-linear amplifier

linear portions, when extrapolated, cut at the so-called 'output intercept point', C. C is measured on the output scale and its value in dB(mW) is a convenient indication of the linearity of the amplifier. From the intercept point, the maximum input an amplifier can handle before the i.p.s. are intolerable can be predicted. Sometimes the 'input intercept point' is referred to, which differs from the output intercept point by the amplifier gain.

An alternative test of linearity is the three-tone test.[2] In this test the vision, colour and sound carriers are simulated by separate tones set at particular levels. These are referred to as f_v, f_c and f_s respectively. The vision tone is set at 8 dB below the reference level of peak carrier power to correspond to an average level of luminance.

In the three-tone test the most visible i.p. appears at (f_v + 1·57)MHz. If the fundamental output represents the peak sync level, the three-tone test i.p.s. at (f_v + 1·57) MHz are shown in Fig. 5. They are 26 dB lower than the two-tone input i.p.s. for the assumed characteristic (see Appendix 4).

For r.b.l. applications an amplifier must be capable of handling four television channels. The maximum amplifier inputs per channel can be expressed in terms of the intercept point to meet the quality requirements of standards P and Q (see Appendix 4). In particular, for four channels, they are:

Standard P: s_i max = $(C-G_\alpha)-24$ dB(mW) (iv)
Standard Q: s_i max = $(C-G_\alpha)-20$ dB(mW) (v)

Incident field strengths are unlikely to exceed 90 dB(μV/m) at sites where pre-amplifiers are considered. This corresponds to an output from four log-periodic aerials of −26 dB(mW) and minimum intercept points of

−2 dB(mW) for Standard P
−6 dB(mW) for Standard Q

2.3. Return loss

The impedance match required at the input and output terminals of a pre-amplifier depends on the level of delayed images which can be tolerated at the final output. These levels are related to the delay times associated with the appropriate signal path loop. When considering the amplifier input impedance the delayed signal is produced by reflection at the amplifier input and re-reflection at the aerial. Similarly when considering the amplifier output impedance the delayed signal results from reflection at the receiver input and re-reflection at the amplifier output.

At low-power relay stations the feeder lengths are short enough to keep the delay time well below 1 μs.

TABLE 2

Amplifier minimum return loss (dB)

v.s.w.r. figures are given in brackets

	Standard P		Standard Q	
	Input Impedance	Output Impedance	Input Impedance	Output Impedance
Mast-head amplifier	12 (1·7)	12 (1·7)	8·5 (2·1)	8·5 (2·1)
Mast-base amplifier	6 (3·0)	18 (1·3)	2·5 (7)	14·5 (1·5)

The delayed signal attenuation required is given in Appendix 3. For a Standard P transmission it is 28 dB and for Standard Q it is 24·5 dB.

If the feeder loss is 3 dB and the return loss at the aerial terminal is 16 dB, the minimum return loss at the amplifier input is given in Table 2 for both mast-head and mast-base situations. Similar figures are given for the minimum return loss at the amplifier output assuming that the minimal receiver input return loss is 10 dB.

2.4. Engelbrecht arrangements

A parallel arrangement of two amplifiers can be used to improve performance.[3] The amplifiers are interconnected by hybrids as shown in Fig. 6: the input signal is divided equally between two paths and fed through separate amplifiers. The output signals are recombined in a second hybrid. The hybrids are designated by two figures which are the output coupling and directivity respectively. Because the reflections can be diverted to a 50Ω termination, the input and output return loss of the system is greater than that of either amplifier alone. The signal path is shared so that each amplifier handles half the input power. This gives the system a 3 dB advantage as far as linearity is concerned. The performance in respect of both noise figure and gain is unchanged. If one amplifier should fail for any reason the output signal is maintained at a 6 dB lower level — hence increased reliability.

3. Assessment of some commercial amplifiers

Altogether 12 amplifiers or arrangements of amplifiers have been examined. Table 3 gives details of the measurements made and also the manufacturers' quoted performance. Amplifiers were chosen either for their low noise figure or high linearity together with low cost. A BBC designed amplifier (reference J) was used as a comparison. The amplifiers' impedance match is generally 75Ω but in an Engelbrecht arrangement designed for 50Ω the mismatch is effectively suppressed.

3.1. Gain

Measurements of gain were made throughout the u.h.f. Bands IV and V* at input levels of about 10 mV, which is well below the overload for these amplifiers. The figures quoted in Table 3 are averages for Bands IV and V with the measured variation over this range.

3.2. Noise

The measured noise factors are the averages taken from a series of spot noise figures measured at 50 MHz intervals in Bands IV and V using a receiver with an i.f. bandwidth of 2 MHz and driven from a source matched to the input impedance of the amplifier. The level of the inherent noise from the pre-amplifier was measured on the receiver. A calibrated noise source was used to exactly double this level.[4] Also tabulated is the comparative noise measure, M, which is given by[5]

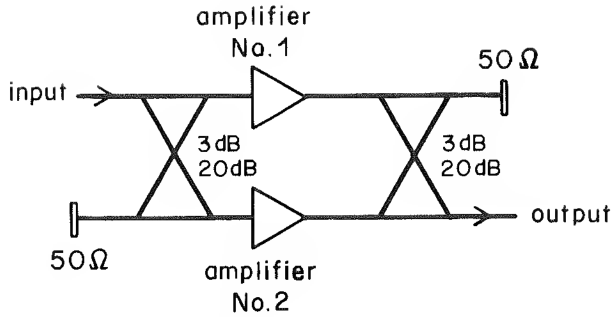

Fig. 6 - An Engelbrecht arrangement of two amplifiers

$$M = \frac{F_\alpha - 1}{1-(1/G\alpha)}$$

where F_α is the noise factor of the amplifier as a power ratio and G_α is the gain of the amplifier as a power ratio. This is a useful noise performance figure of merit because it takes account of the amplifiers' gain. The lower noise measure indicates the better amplifier.

3.3. Linearity

The output and input intercept points are determined from two-tone test measurements at 650 MHz. From these the maximum level for a four-channel input, assuming a worst i.p. of 46 dB relative to the output, can be calculated (see Appendix 4). Manufacturers normally state the linearity performance of amplifiers in terms of the German specification DIN45004.[6] The quoted figures in Table 2 have been derived from this.

3.4. Impedance match

The input and output v.s.w.r.'s are given referred to the nominal characteristic impedance. This is 75Ω except for the Engelbrecht arrangements and the BBC designed amplifier (J) which is 50 ohms.

4. Improvement of amplifier noise performance

The noise factor is defined as the ratio of

(i) the total noise power per unit bandwidth available at the output port when the noise temperature of

* Band IV 470–582 MHz
Band V 614–854 MHz

TABLE 3

The performance of some commercial amplifiers

Model	Power supply	Gain (dB) in Bands IV and V		Noise			Linearity				Impedance match	
				Noise factor (dB)		Noise § measure (dB)	Output intercept dB(mW)	Input intercept dB(mW)	Max. input dB(mW) #		Input	Output
		Quoted	Measured	Quoted	Measured				Quoted	Measured	v.s.w.r.	v.s.w.r.
A	24V 34mA	18	16 ± 1	5	4·5 ± 0·5	3·49	21·5	+4·5	–15·5	–18·5	2·1	2·45
A ENGELBRECHT	24V 68mA	18	16 ± 2	5	5 ± 0·5	4·24	24·6	+6·6		–16·4	1·95	1·77
B	24V 24mA	22	20 ± 2	5	5 ± 0·5	4·13	27·8	+3·8	–27·5	–19·2	2·60	2·25
C	16V 3mA	16	12 ± 2	3·5	5 ± 0·5	4·52	14·3	+0·3	–24·5	–22·7	–	–
D	24V 34mA	22	20 ± 1	3·5	6 ± 1	6·09	24·3	+3·3	–16·5	–19·7	7·2	2·60
D ENGELBRECHT	24V 68mA	22	20 ± 1·5	3·5	6·5 ± 0·5	6·10	28·3	+6·8		–16·2	1·69	1·74
E *	INTERNAL	28	27 ± 2	4·5	6 ± 0·5	5·41	32·0	+3·0	–17·9	–20·0	3·1	2·0
F †	24V	15	14 ± 2	7	7·5 ± 0·5	7·53	23·7	+7·7	–13·8	–15·3	1·18	1·21
G †	24V	25	26 ± 4	5·5	7·5 ± 1	7·87	20·6	–9·4	–28·8	–32·4	1·55	1·70
H	INTERNAL	20 ± 1	22 ± 1·5	6·5–10	8·5 ± 1	9·02	31·0	+7·5	–22·5	–15·5	1·77	2·33
I	24V 20mA	16 ± 2	14 ± 1	7	5·5 ± 1	5·62	25·0	+10·0	–25·5	–13·0	3·55	4·15
J	–12V 80mA	21 ± 3	22 ± 2	7	9·0 ± 0·5	9·03	23·0	–1·0	–27·0	–24·0	2·20	1·86

† Engelbrecht arrangement § See text # for 4 u.h.f. ch. 46 dB i.p.'s * VHF also

the input termination is standard (290°K), to

(ii) that portion of (i) engendered by the input termination.

The spot noise factor is that measured in a narrow frequency band and is useful for specifying the noise factor at individual frequencies. As defined the noise factor depends on the internal structure of the amplifier and on its input termination — but not on its output termination, i.e.

$$F = f(F_o, Y_s, i_E)$$

where F_o represents the contribution from the amplifier itself, Y_s is the source admittance and i_E is the emitter current of the first transistor stage. There exists, therefore, an "optimum" noise factor.[1,7,8] This occurs at a particular frequency and is the lowest noise factor which can be obtained through adjustment of the source admittance and transistor operating point.

For amplifier A an attempt was made to improve its noise performance by adjusting the source admittance. By doing this for two amplifiers in an Engelbrecht configuration the overall impedance match would remain acceptable. Fig. 7 shows how the noise factor of a single amplifier varies with frequency under (a) matched conditions and (b) driven from a 50Ω source. Further a microstrip matching network was constructed consisting of a 50Ω line with three shunt variable capacitors at eighth wavelength intervals. The best noise performance which was found for a particular source impedance maintained across the frequency band is shown in Fig. 7 (c). It is only marginally better than the matched case and that only over a limited frequency range.

Also, for the same amplifier A, the operating point was effectively changed by reducing the supply voltage — but this also reduces the gain. The result of a 10V supply voltage is shown in Fig. 7 (d).

To achieve a good all-round performance it is important to have the correct balance of noise factor, gain and linearity.

5. An RBL system using inexpensive amplifiers

The proposed r.b.l. front-end shown in Fig. 8 comprises a four-channel bandpass mast-head filter, a mast-head amplifier, a main feeder and an active splitter. The filter is required to protect the pre-amplifier from out-of-band signals and from the local transmission. The mast-head filter and pre-amplifier require a weatherproof box for mounting and are expected to work to a high degree of reliability. The active splitter is a wideband device which provides four separate outputs with an adequate degree of isolation between them.

The filter, with a bandwidth of about 100 MHz to

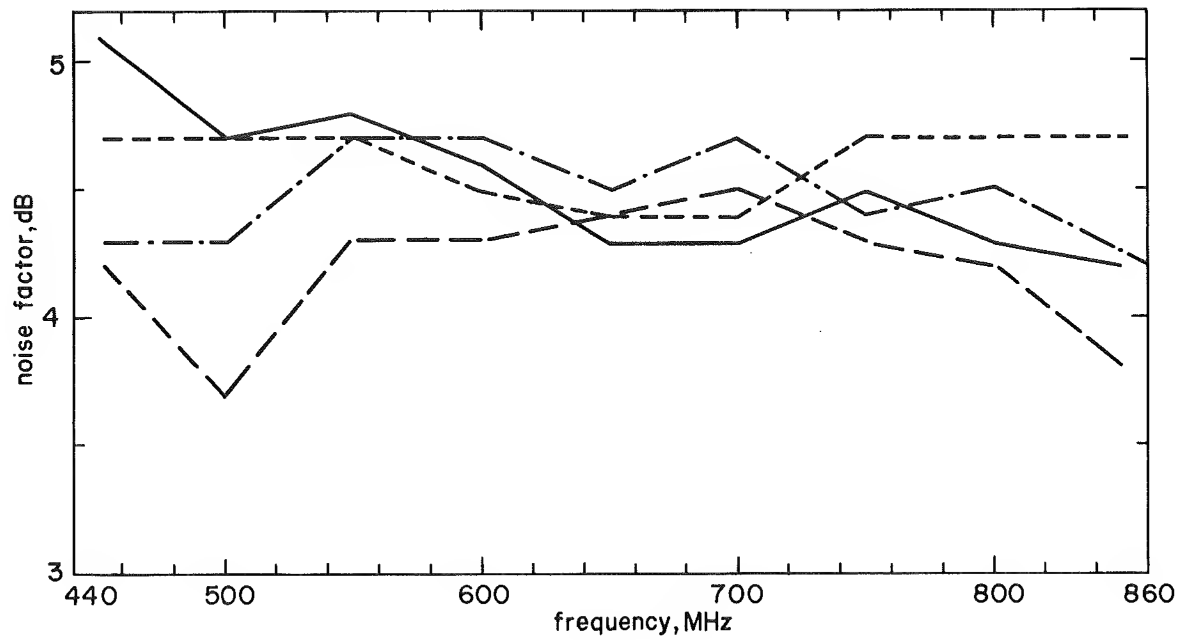

Fig. 7 - The noise factor of amplifier A under different source impedance conditions

(a) 75Ω source (b) 50Ω source (c) optimum source (d) 10V supply volts

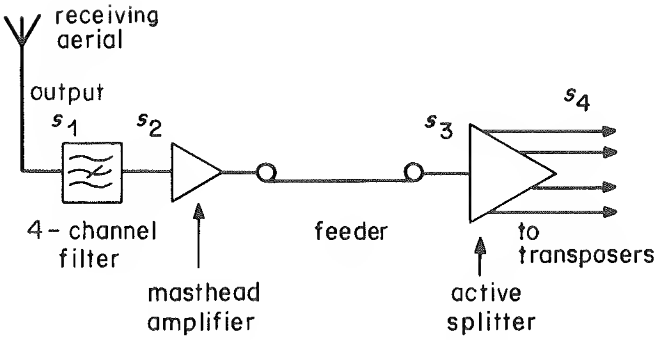

Fig. 8 - An RBL system using inexpensive amplifiers

s_1, s_2, s_3 and s_4 refer to the power levels at the respective points

cover a standard four channel grouping, must be of low loss, say less than 1 dB, to avoid degrading the overall low-signal performance. Suitable filters are commercially available: the band-edge slope is dependent on the number of sections chosen for the filter.

From the amplifiers examined, a suitable choice has been made to fit into the general arrangement of Fig. 8. The mast-head amplifier must be of the low-noise type — an Engelbrecht pair using amplifier A units would be appropriate. For the wideband splitter the overload performance is more important than the noise performance. Its gain need only be sufficient to overcome the losses associated with the four-way split. A suitable choice would be amplifier module units F in Engelbrecht form. The splitting arrangement consists of three hybrids and divides the signal in two stages; this is described in more detail in the next section. Typical operating levels for this arrangement are indicated in Table 4 for 650 MHz.

The assumptions are that

Mast-head filter loss	=	1 dB
Pre-amplifier gain	=	16 dB
RBL feeder loss	=	4 dB
Active splitter, insertion gain to each port	=	8 dB

Three different receiving aerials are considered, comprising one, two or four log-periodic aerials (l.p.'s). Transposers of 2W and 10W are considered separately and the minimal and maximal levels given are based on typical transposer operating levels. The minimal levels are set by signal to noise ratio limits and the maximal levels by inter-channel cross-modulation, assuming a standard P transmission.

6. An active 4-way splitting unit

A 4-way splitting unit has been constructed from an Engelbrecht pair of amplifier F units and five hybrids made from wireline and interconnected by microstrip on double-sided printed laminate — see Fig. 9.

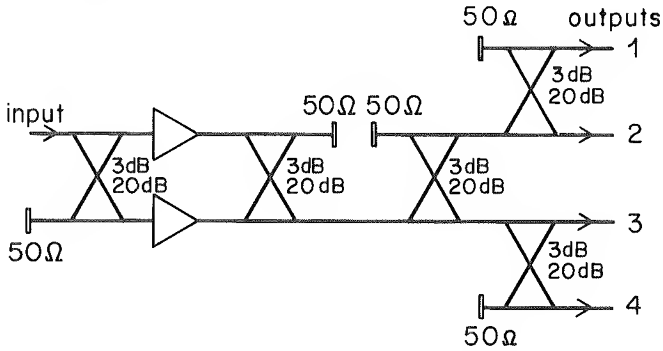

Fig. 9 - An active 4-way splitting unit

The high- and low-signal performance of amplifier F has already been referred to in Table 3. In this section the results of measurements are presented. Firstly, the insertion gain of the splitting unit in Bands IV and V is shown in Fig. 10.

Secondly, the isolation between the output ports is given in Fig. 11: it is better than 15 dB between any two ports.

Thirdly, the worst return losses at the output ports are

Output port number	Worst return loss (dB)
1	22
2	20
3	20
4	24·5

TABLE 4

Transposer power		Received field strength dB(μV/m) — Receiving aerial and gain (dB): 1 l.p. 8 dB	2 l.p. 11 dB	4 l.p. 12·5 dB	Aerial output s_1 dB(mW)	Pre-amp input s_2 dB(mW)	Active splitter input s_3 dB(mW)	Transposer input s_4 dB(mW)
2W	min	69·9	66·9	65·4	–53·4	–54·4	–42·4	–34·4
	max	92·5	89·5	88·0	–30·8	–31·8	–19·8	–11·8
10W	min	69·9	66·9	65·4	–53·4	–54·4	–42·4	–34·4
	max	89·3	86·3	84·8	–34	–35	–23	–15

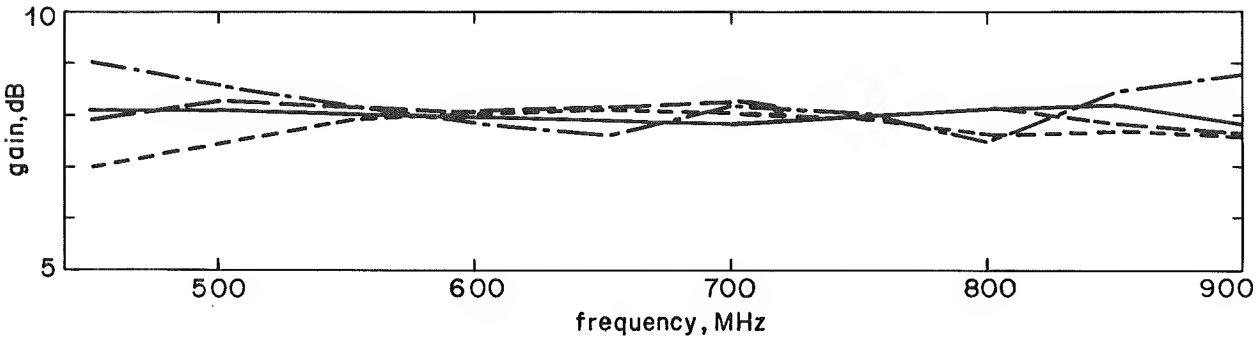

Fig. 10 - Gain of the active splitter using wireline hybrids

———— input — port 1 - - - - - - - input — port 2 — — — input — port 3
— — input — port 4

7. Aerial isolation and input filtering

This section discusses the isolation which is required between the terminals of the transmitting and receiving aerials at a low-power relay station with regard to the generation of unwanted inter-modulation products. With up to four television channels amplified in a common amplifier, the worst of these take the form of spectral components near to the vision carrier frequency. Subjectively this appears as a patterning similar to that produced by some forms of co-channel interference. In general closed-loop effects such as system stability and the visibility of delayed images created in the loop, will be insignificant if the i.p. criteria are met.

In principle, any desired loop transmission loss may be achieved by inserting a channel-group filter before the mast-head amplifier. The specification for this filter is derived. Maximal acceptable i.p. levels are derived from the limits for a standard P transmission (see Appendix 3).

It is convenient to consider the example of the r.b.l. system described in Section 5. Assuming that the losses associated with the main feeder and distribution transformers etc. are 4 dB, the transmitting aerial input levels are

For 2W transposer: +29 dB(mW)

For 10W transposer: +36 dB(mW)

If I_A is the isolation between the terminals of the transmitting and receiving aerials, the local transmission appears at the r.b.l. aerial terminal at a level of

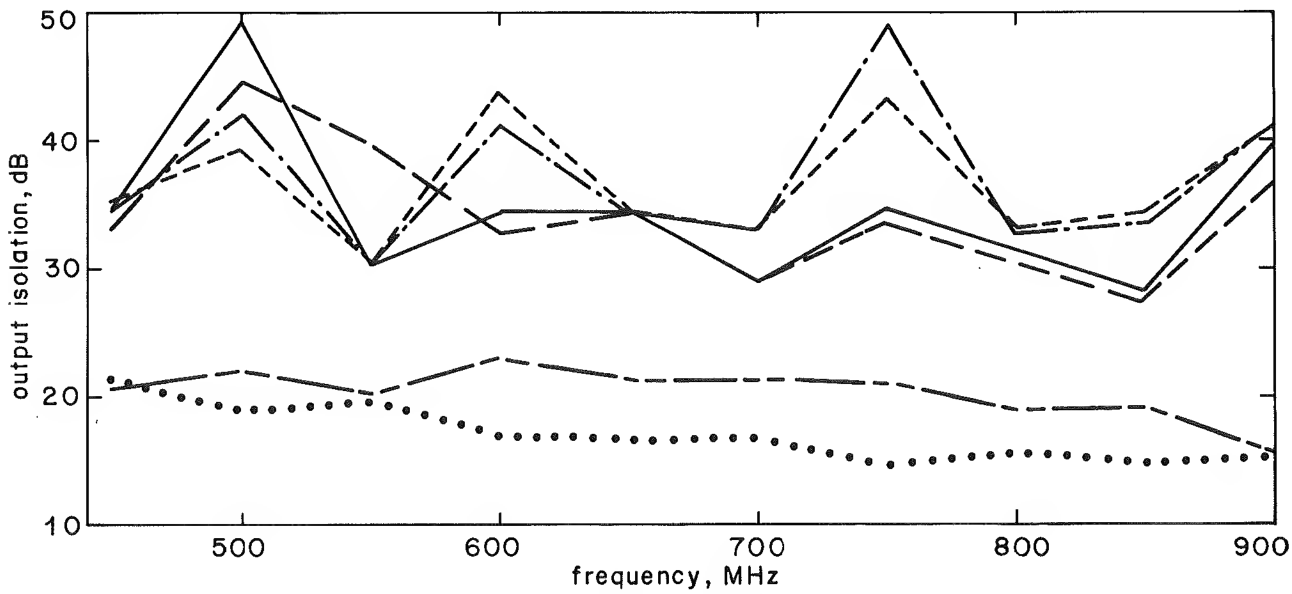

Fig. 11 - Output isolation of the active splitter using wireline hybrids

———— ports 1–4 - - - - - - - ports 2–4 — — — ports 2–3
— — ports 1–3 — — — ports 3–4 ••••••••• ports 1–2

For 2W transposer: $(29-I_A)$ dB(mW))

For 10W transposer: $(36-I_A)$ dB(mW)) (vi)

The maximum permissible level of this unwanted input is set by the linearity considerations. One of the worst situations for a u.h.f. standard channel group transposition is shown in Fig. 12. This shows the spectrum of vision carriers present at the receiving aerial terminal. The peak sync level of the wanted signal in a Lower Band V channel group is represented by s_1. The peak sync level of the unwanted signal in an Upper Band V channel group is represented by s_2. It is assumed that without a mast-head filter the local transmission may be a stronger signal than the wanted.

The worst channel i.p. in this situation is that in Channel 51 at the vision carrier frequency which is produced by the vision carriers of channels 55 and 59 when they reach peak sync level. The level of this i.p. at the transposer input is given by s_{o2} (see Equation xvi, Appendix 4).

$$s_{o2} = 3(s_2 + G_\alpha) - 2C \text{ dB(mW)}$$

Where G_α is the r.b.l. system gain and C is the output intercept point.

The wanted signal level at the transposer input is $(s_1 + G_\alpha)$. The relative i.p. level is given by s'_{o2} where

$$s'_{o2} = s_1 + 2(C-G_\alpha) - 3s_2 \text{ dB} \qquad \text{(vii)}$$

and $(C-G_\alpha)$ is the input intercept point which in this example is $-4{\cdot}3$ dB(mW).

The i.p. performance data applicable to a standard P transmission is given as -48 dB in a 3-tone test. This implies a single channel only. When more channels are amplified simultaneously in a common amplifier, the cross-modulation of the vision carriers is significant and the protection figure of -55 dB for co-channel interference is more appropriate.

From equation (vii) the maximum unwanted input level is found by assuming that the lowest wanted input

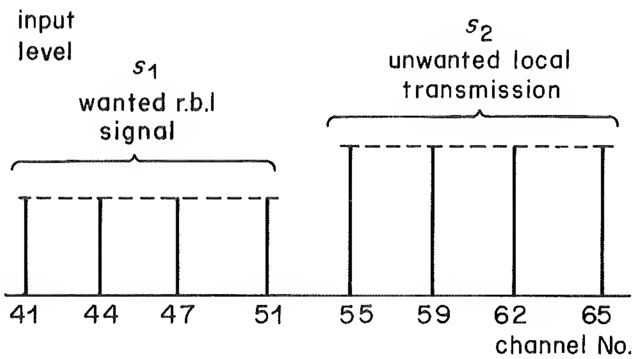

Fig. 12 - The signal spectrum at the input of an RBL system

level is $-54{\cdot}4$ dB(mW)

$$\therefore s_{2max} = \frac{1}{3}\left[s_1 + 2(C-G_\alpha) - s'_{o2}\right]$$

$$= -39{\cdot}3 \text{ dB(mW)} \qquad \text{(viii)}$$

The minimum aerial isolation with no mast-head filter is seen to be (referring to (vi) and (viii) as follows:

for 2W transposer: $29 + 39{\cdot}3 = 68{\cdot}3$ dB,

for 10W transposer: $36 + 39{\cdot}3 = 75{\cdot}3$ dB.

The addition of a mast-head filter allows the aerial isolation to be relaxed further. Consider the effect of a channel group filter such that the response at the band edge is as shown in Fig. 13.

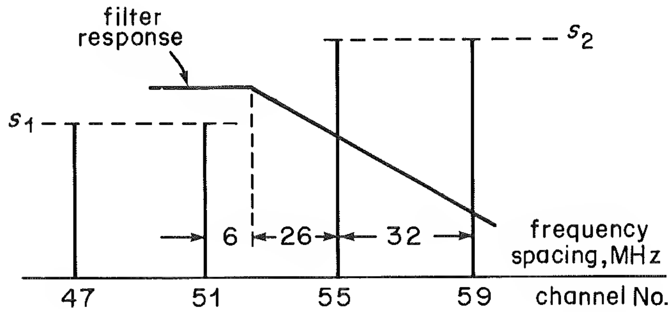

Fig. 13 - Filter response at band-edge

The level of the unwanted signal at the r.b.l. aerial terminal is reduced by the slope of the filter response. If the response falls off at xdB/MHz, the i.p. amplitude is reduced by a total of

$$(2 \times 26)x + 58x = 110x \text{ dB} \qquad \text{(ix)}$$

Assuming that the filter insertion loss is 1 dB, the unwanted signal level is given by:

for 2W transposer: $(28-I_A) - \frac{110}{3} x$ dB

for 10W transposer: $(35-I_A) - \frac{110}{3} x$ dB

The maximum unwanted signal level is given by (viii). We have, therefore, a numerical relationship between the aerial isolation and the filter response.

For 2W transposer:

$$(28-I_A) - \frac{110}{3} x = -39{\cdot}3 \text{ dB}$$

$$\therefore 3I_A + 110x = 202 \text{ dB} \qquad \text{(x)}$$

For 10W transposer:

$$(35-I_A) - \frac{110}{3}x = -39{\cdot}3 \text{ dB}$$

$$\therefore \; 3I_A + 110x = 223 \text{ dB} \qquad \text{(xi)}$$

These are shown graphically in Fig. 14.

8. Measurements of aerial isolation

Attention is directed to the low-power u.h.f. stations and the aerial arrangements representative of these installations. These will usually be log-periodic aerials, although up to two panel aerials or a simple cardioid aerial are alternatives. A series of experiments is described in this section to evaluate the likely aerial isolation which it will be possible to achieve at these stations.

A simulated mast, consisting of a main triangular-section of braced steel extended at one end by a circular section, of diameter similar to a cantilever pole, is supported horizontally about 1·6m above the ground. The arrangement is illustrated in Fig. 15. The receiving aerials are mounted on a trolley which can be moved along the top edge of the triangular section to vary the physical separation between the aerials. The separation is measured to the first transmitting aerial element. The isolation is determined by substituting a calibrated attenuator for the aerial path and maintaining the signal level constant.

Measurements were made initially at several frequencies in Bands IV and V and a periodic variation with frequency was noticed at a given separation. A similar variation is apparent as the separation is varied at one frequency. Since it is the worst case situation which is of interest here, all subsequent measurements were made at 700 MHz, which frequency was affected little by local off-air transmissions.

Two templets indicating the minimum aerial isolation are produced in Fig. 16. One is for simple arrays of log-periodic aerials, i.e. a single, double or 4-tiers of one aerial per tier in the transmitting case and one or two aerials in the receiving case. The second templet covers more complicated transmitting arrays with 2 aerials/tier. Measurements were made with the included angle between transmitting and receiving aerials set at 0°, 45° and 90° and with the receiving aerial on either polarisation. A greater isolation would be expected if four aerials were used in the receiving array.

The results show that, for a large number of cases, the minimum aerial separation will be 3·8m at channel 21 and 2·2m at channel 68. These separations assume that there is no significant contribution to the isolation from the channel filter.

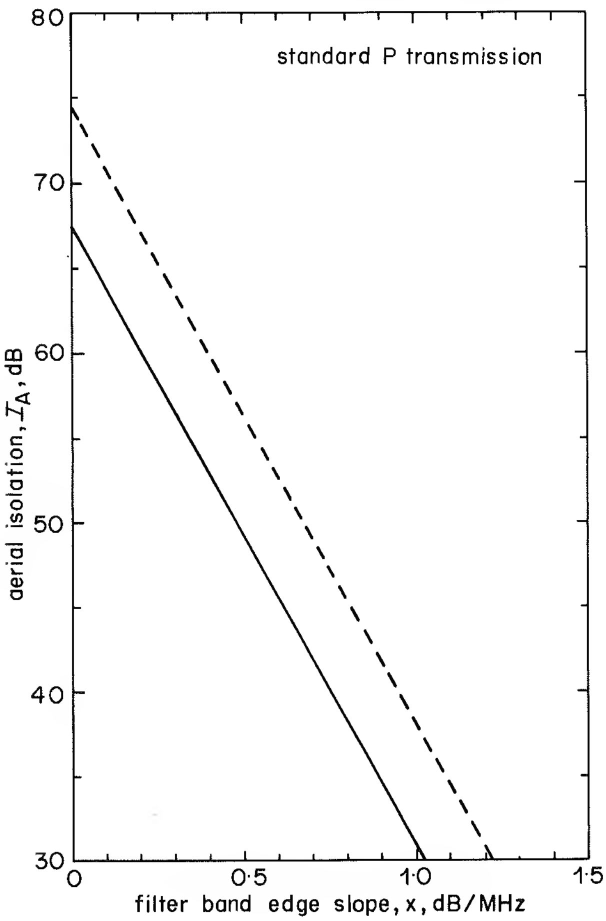

Fig. 14 - The interrelation of aerial isolation and mast-head filter response in an RBL situation

——— 2W transposer
- - - - - 10W transposer

9. Conclusions and recommendations

There are two main advantages of using pre-amplifiers at low-power u.h.f. relay stations:

a) where co-channel interference conditions allow, the number of stations which require only a single or double log-periodic receiving aerial may be increased;

b) a useful increase in the sensitivity of the receiving system may be achieved; thus, by using a mast-head pre-amplifier, the received field strength may be 9 dB lower than when a pre-amplifier is not used. The corresponding figure for a mast-base amplifier is 6 dB.

Some inexpensive amplifiers have a performance in respect of noise figure and linearity comparable with that of more expensive professional units. Their impedance levels are usually specified to be 75Ω but an acceptable match to 50Ω can be achieved in an Engelbrecht configuration. It is therefore recommended that the use of amplifiers of this type should be considered in the design of receiving arrangements at u.h.f. relay stations.

Critical matching at the input terminals of the pre-amplifiers was found to achieve a marginal improvement

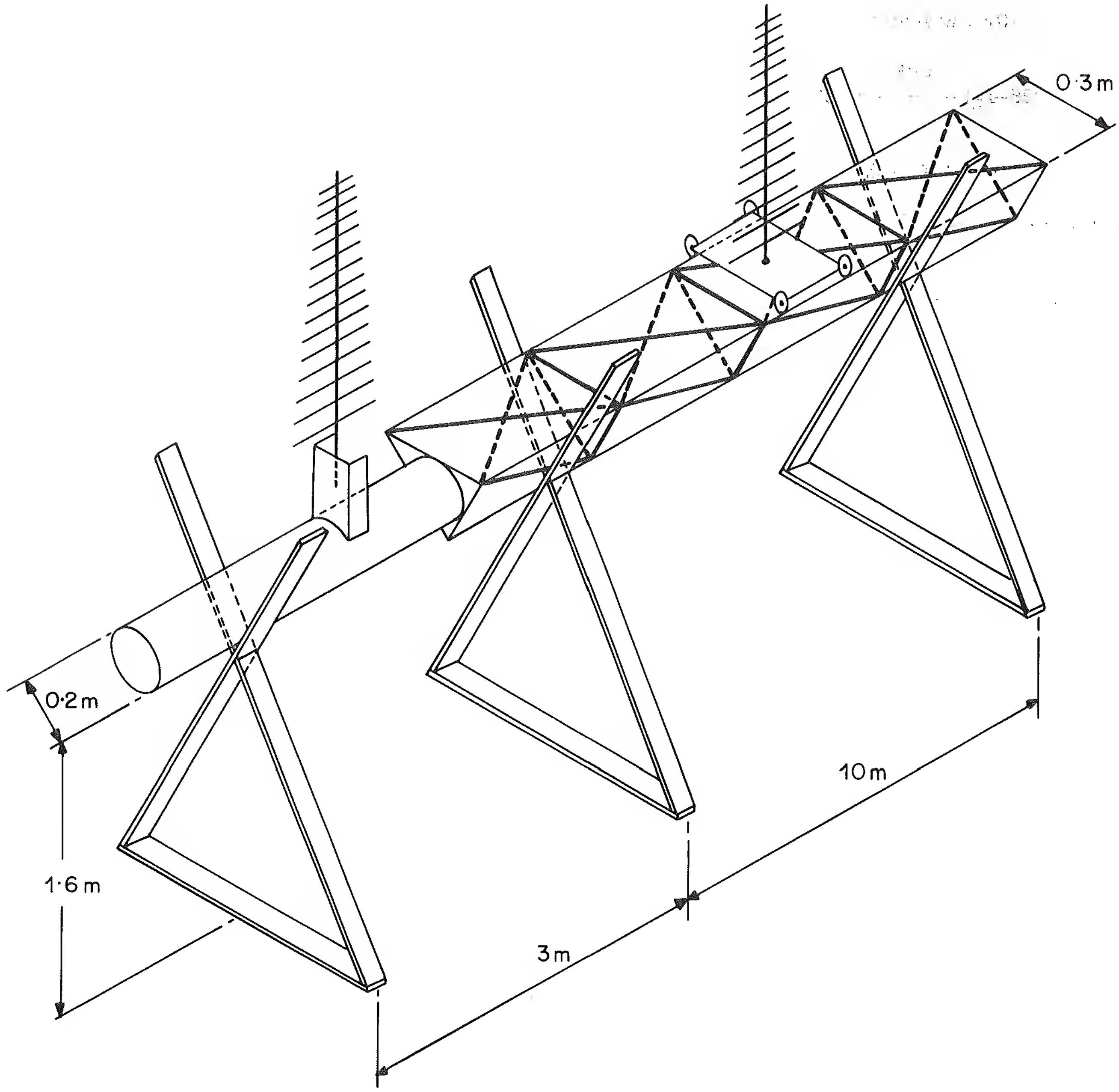

Fig. 15 - Measurement of aerial isolation

only and is not considered to be a useful technique without amplifier re-design.

An active four-way splitter has been described which can replace the splitting filters which have been used to date. It may be possible to produce a cheaper version using microstrip techniques.

A mast-head bandpass filter is desirable for protection from strong out-of-band signals. The protection required from the local transmissions is less stringent. The band-edge slope of the filter requires only a few sections to meet the specification.

Two templets of aerial isolation have been produced – one for simple transmitting arrays comprising a single aerial per tier and a second for more complicated arrays with two aerials per tier.

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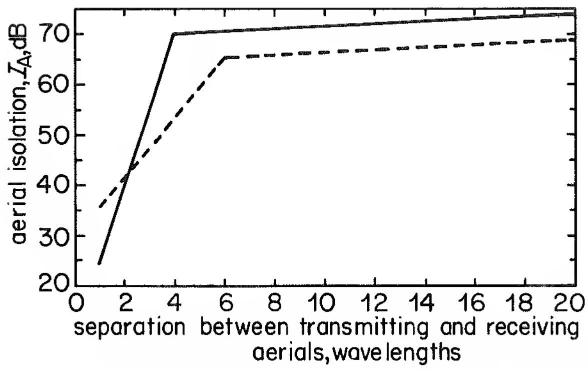

Fig. 16 - The isolation between the RBL and transmitting aerials for RBL installations

——————— arrays with single aerial/tier
– – – – – – – arrays with two aerials/tier

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APPENDIX 1

The power available across the matched load of an aerial

The equivalent circuit of a receiving aerial and load is shown in Fig. 17. Under matched conditions, the voltage developed across the load is given by:

$$v_L = 20 \log \frac{\lambda}{2\pi} + E + G_A \qquad \text{dB}(\mu\text{V})$$

where G_A is the aerial gain relative to $\frac{\lambda}{2}$ dipole, dB

E is the received field strength, dB(μV/m)

λ is the wavelength in metres

The power in the matched load is numerically related to v_L by

$$W_L = v_L - 108{\cdot}6 \qquad \text{dB(mW)}$$

$$\therefore \quad \underline{W_L = E + 20 \log \lambda + G_A - 124{\cdot}6 \qquad \text{dB(mW)}} \qquad \text{(xii)}$$

For the u.h.f. range:

Freq. (MHz)	20 log λ
470	−3·9
600	−6·0
650	−6·7
720	−7·6
860	−9·1

The net aerial gain including feeder loss G_A may be assumed to take the following values for log-periodic (l.p.) arrays

1	l.p.	:	8 dB
2	l.p.s.	:	11 dB
4	l.p.s.	:	12·5 dB

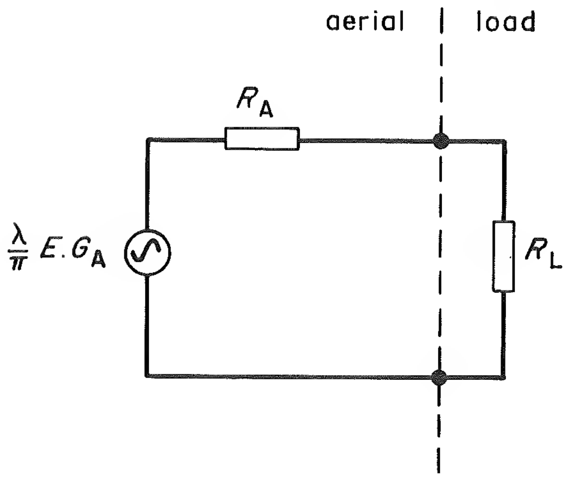

Fig. 17 - Equivalent circuit of a receiving aerial and load

APPENDIX 2

The signal to noise ratio of a receiving system

The unweighted signal-to-noise ratio of a base-band television signal is defined as

$$\frac{S}{N} = 20\log_{10}\left(\frac{\text{pk-pk picture voltage}}{\text{rms noise voltage}}\right) \text{ dB}$$

Assuming that the pk-pk picture voltage is 56% (–5 dB) of the peak sync.carrier voltage at the receiver input, the pk-pk picture voltage is given by

$$\text{pk-pk picture voltage} = \text{rms of peak sync.carrier level} -5 +3 \quad \text{dB}$$

$$\therefore \quad \frac{S}{N}\text{ at receiver input} = 20\log_{10}\left(\frac{\text{rms of peak sync.carrier voltage}}{\text{rms noise voltage}}\right) -2 \quad \text{dB}$$

$$= 10\log_{10}\left(\frac{\text{peak sync.carrier power}}{\text{rms noise power}}\right) -2 \quad \text{dB}$$

Also,

$$\frac{S}{N}\text{ at receiver output} = \frac{S}{N}\text{ at input} - F_s - 6 \quad \text{dB}$$

where F_s is the noise factor of the receiver and the reduction of 6 dB accounts for the receiver response.[9]

$$\therefore \quad \frac{S}{N}\text{ at receiver output} = 10\log_{10}\left(\frac{\text{peak sync.carrier power at input}}{\text{rms noise power at input}}\right) - F_s - 8 \quad \text{dB}$$

The peak sync.carrier power available at the input of a receiver matched to the aerial is given by (xii). The noise power associated with an aerial, assumed to be at a temperature, T, of 290°K. in an effective bandwidth, B, of 5·08 MHz is given by

$$N = \mathrm{K}TB = -106{\cdot}9 \quad \text{dB(mW)}$$

where K is Boltzman's constant

$$\therefore \quad \underline{\frac{S}{N}\text{ at receiver output} = E + 20\log\lambda + G_A - F_s - 25{\cdot}7 \quad \text{dB}} \qquad \text{(xiii)}$$

where E is in dB(μV/m)

G_A, F_s in dB

APPENDIX 3

Relevant minimum transmission standards at the output of UHF relay stations

Impairment		Standard P	Standard Q
Signal-noise ratio		39 dB	33 dB
Attenuation of delayed images:			
$\leqslant 1 \mu s$ delay		28 dB	24·5 dB
$> 1 \mu s$ delay		30·5 dB	28 dB
Intermodulation products (level of 1·57 MHz i.p. in a 3-tone test)		–48 dB	–40 dB
Co-channel interference			
zero offset:	99% time	55 dB	50 dB
	1% time	45 dB	40 dB
$+\frac{5}{3}$ offset:	99% time	45 dB	40 dB
	1% time	30 dB	30 dB

APPENDIX 4

Amplifier linearity

Let the amplifier characteristic be represented by a third-order polynomial such that the output response, v_o, to an input signal v_i, is

$$v_o = \alpha v_i + \beta v_i^3$$

where α and β are constants.

When the input takes the form of a dual tone, $v_o = V_i \cos \omega_r t + V_i \cos w_s t$, the amplitude of the i.p.s., v_{o2}, is $\frac{3}{4}\beta . V_i^3$. The power level associated with these i.p.s. is given by

$$s_{o2} = -2{\cdot}5 + \beta + 3s_i \quad \text{dB(mW)}$$

where s_i is the input power level.

Interpreting from Fig. 5, we have

$$C = \tfrac{1}{2}(3\, G_\alpha - \beta + 2{\cdot}5) \quad \text{dB(mW)} \qquad \text{(xiv)}$$

$$\beta = 3\, G_\alpha + 2{\cdot}5 - 2C \quad \text{dB} \qquad \text{(xv)}$$

and $$s_{o2} = 3(s_i + G_\alpha) - 2C \quad \text{dB(mW)} \qquad \text{(xvi)}$$

Note that the constant α is seen to be the amplifier gain and, is therefore written G_α.

Further if the input has three components with appropriately chosen frequencies, a television channel can be simulated. Let,

$$v_i = V_v \cos \omega_v t + V_c \cos \omega_c t + V_s \cos \omega_s t$$

where ω_v, ω_c and ω_s represent the vision, colour and sound carriers and V_v, V_c and V_s represent the levels of the carriers respectively. The amplitude of the i.p. at $(f_v + 1{\cdot}57)$MHz is $\frac{3}{2}\beta\, V_v V_c V_s$. The power level of this i.p. is given by

$$s_{o3} = +3{\cdot}5 + \beta + s_v + s_c + s_s \quad \text{dB(mW)}$$

where s_v, s_c and s_s are the power levels of the vision, colour and sound components respectively.

If s_i is the input level corresponding to peak syncs,

$$s_{o3} = \beta + 3s_i - 28{\cdot}5 \quad \text{dB(mW)}$$

and from (xv)

$$s_{o2} = 3(s_i + G_\alpha) - 2C - 26 \quad \text{dB(mW)} \qquad \text{(xvii)}$$

which referring to (xvi) is 26 dB lower than s_{o2}.

Expressed relative to the output, the i.p. level is given by,

$$s'_{o3} = (s_i + G_\alpha) - 3(s_i + G_\alpha) - 2C - 26 \quad \text{dB}$$

$$\therefore \quad s'_{o3} = 26 + 2C - 2G_\alpha - 2s_i \quad \text{dB} \qquad \text{(xviii)}$$

Re-arranging, we have an expression for the maximum permitted input level for the amplifier,

$$s_i \max = \tfrac{1}{2}(26 + 2C - 2G_\alpha - s'_{o3}) \quad \text{dB(mW)} \qquad \text{(xix)}$$

A computer program has been used to evaluate the effect of further channels. The worst i.p. amplitude is found and from the results the maximum input is given by:

$$\underline{s_i \max = (C - G_\alpha) - \mathrm{Z} \quad \text{dB(mW)}} \qquad \text{(xx)}$$

where $(C–G_\alpha)$ is the input intercept point and Z is given in Table 5.

TABLE 5

	Relative i.p. level (dB)						
No. of channels	40	46	48	50	52	55	60
1	7	10	11	12	13	14·5	17
2	16	19	20	21	22	23·5	26
4*	20	23	24	25	26	27·5	30

HR/VY

* Channels are assumed to be not equispaced, e.g. 3—3—4 arrangement

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